DEVELOPMENT OF BROADBAND MICROWAVE COUPLER USING MICROSTRIP - SLOT LINES IN MIC ENVIRONMENT

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CERTIFICATE

It is certified that the work contained in the thesis entitled **Development of broad band microwave coupler using microstrip slot lines in MIC environment"** by **Sukhbir Singh,** has been carried out under my supervision and that this work has not been submitted elsewhere for PGDIIT

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Abstract

A novel method of realization of microstrip slot coupler with compensation is presented in this thesis Generally microstrip couplers are realized using quarter wave planar transmission lines This type of microwave coupler uses both sides of the substrate one side of which consists of microstrip line and the other side consists of a slotline The substrate required for this coupler should have high permittivity due to disperssive nature of the slotline Alignment for this type of coupler should be done in such a way that slotline must lie just beneath the microstrip line on the other side of the substrate Analysis with and without compensation of the same coupler is made in this thesis by considering double symmetry of the reciprocal four port network Compensation to the coupler is given by extending the length of slotline symmetrically on both sides of the terminating ends Scattering parameters are derived for both the compensated and uncompensated cases of the coupler structure The realization of the coupler with design parameters corresponding to 4 5 dB coupler on the alumina substrate with relative permittivity 9 8 and substrate thickness 0 635 mm is implemented over RT/ Duroid substrate (ε_r=10 2) with thickness 0 325 mm For this substrate approximate calculations are made using empirical formulae Approximated theoretical center frequency is achieved around 6 7404 GHz Measurements are taken over the band of interest (from 6 0 GHz to 8 0 GHz) The individual reponses of the implemented coupler are measured and compared with the theoretical ones

To my parents

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CHAPTER 1

1 1 INTRODUCTION

Hybrids and couplers are indispensable components in rapidly growing applications of microwaves in electronics warfare radars and communication systems The circuits are used fiequently in frequency discriminators balanced amplifiers balanced mixers automatic level control and many other applications a hybrid or a coupler can in piinciple be represented as a multi port network In the case of four port network the four ports of structure are input direct coupled and isolated The two most important parameters that describe the performance of the network are its coupling factor and directivity

A large number of papers on microwave directional couplers have appeared during the past 50 years Some have reported advances in existing types while others have described entirely new types or classes of directional couplers The microwave Engineer is faced with a situation whereby he is frequently able to satisfy a specification by the use of one of several types of direction couplers The choice is then made on the basis of cost manufacturing techniques and the physical layout etc Quite as often the specification may be met only one type of couplei so that knowledge of the entire field is desirable

1 2 Classification of couplers

Attempts are often made to classify directional couplers in various ways In Practice it is found that no unique classification is possible A classification may be based on one of many sets of conditions and the particular sets in one scheme rarely coincide with those in another Such schemes are given as following

1 2 1 Co- or Contra directional coupling between parallel Waveguides or Transmission Lines In a co directional coupler the coupled wave travels in the

same direction as that of the wave in the main line while in the contra directional coupler the coupled wave travels in the opposite direction i e is coupled backward

1 2 2 Interference or intrinsic coupling, This applies to coupling between waveguide or transmission lines by means of longitudinal coupling (discrete or continuous) In an interference coupler the individual apertures or elemental parts of continuos aperture are not themselves directional in their coupling but the individual coupled waves tend to reinforce in one direction of propagation and to interfere destructively in the other In the case of intrinsic coupling the individual apertures are themselves directional in their coupling properties In this case the apertures may be cascaded to give an additional interference which enhances the directivity

1 2 3 Classification by phase division Most types of directional couplers may be classified as either 90 degree (quadrature) couplers or as 180 degree (magic Tee type) of couplers at least at their midband frequency They usually retain their 90degree or 180 degree property within some tolerance over the operating band but some asymmetric) types do not

1 2 4 Classification by the form of coupling There are many ways in which coupling may be achieved some of the various methods are single aperture coupling multi aperture coupling continuous (longitudinal) aperture coupling branch line or guide coupling waveguide junctions ring couplers and quasi optical couplers

1 2 5 Rectangular Waveguide In the case of rectangular waveguide directional couplers the coupling is often classified as either narrow to narrow wall broad to broad wall or as narrow to broad wall The first two methods are most common and are called simple broad wall or narrow wall couplers respectively

This merely states that the types of guiding System in which the directional couplers may be designed e g rectangular ridged or circular waveguide coaxial and strip transmission lines quasi optical guides etc In general the fact is this that nearly all

Couplers may be analyzed in terms of the even and odd normal modes which means that the mathematical analysis of all these types are all much the same

We are interested in the implementation of directional coupler of which design involves planar transmission lines Because this type of transmission structure is suitable in microwave integrated circuits A planar geometry implies that the characteristics of the element can be determined from the dimensions in a single plane

1 3 Classification of transmission lines

Various forms of planar transmission lines have been developed for use in microwave integrated circuits [18] The strip line microstrip line inverted micro strip line slot line co planar waveguide and co planar stripline are respectively planar transmission lines The circuits realized using any one of the aforementioned transmission lines or combinations of them have distinct advantages compared to conventional microwave circuits such as lightweight small size improved performance better reliability and reproducibility and low cost They are also compatible with solid state chip device Above mentioned planar transmission lines description is as follows

1 3 1 Striplines Strip line is one of the most commonly used transmission lines for passive MICs The dominant mode in a stripline is TEM

Microstrip Lines Unlike the stripline the microstrip line is an inhomogeneous transmission line [15] since the field lines between the strip and ground plane are not contained entirely in the substrate Therefore the mode propagating along the microstrip is not purely TEM but quasi TEM The microstripline has become the most widely used transmission line medium because of its unique distinguishing quality needed by planar circuits Its unique characteristics include a highly compact and rugged structure Large scale integration feasibility and ease of device and component implementation To ensure single mode propagation both strip width (w) and substrate thickness (h) are

usually held to a small fiactıon of the operatıng wavelength At low fiequency where quası TEM pıopagatıon ıs assumed the mıcrostrıp ıs usually open structured However at mıllımeteı wave fiequencıes the ınıcrostııp lıne generally requıres metallıc shıeldıng foı radıatıon ınteıference And hıgheı order mode excıtatıon etc There wıll be pıesence of waveguıde modes ın the shıelded structuıe To suppress waveguıde mode propagatıon and allow only the quası TEM mode to propagate along the lıne channel must be desıgned so that the cut off frequency of domınant waveguıde mode whıch ıs generally hybııd mode as the cuııents ın the stııp flow both the longıtudınal and transverse dırectıons Cut off of the domınant waveguıde mode also depends on cross sectıon of the shıeldıng stıuctuıe In case of paıallel coupled mıcrostrıps most of the E field lınes between mıcıostrıps are paıallel to ground plane whereas ın odd mode case most of the E field lınes between strıps are perpendıcular to gıound plane If the structure ıs pure TEM both odd and even mode phase velocıtıes wıll be same Here as strıps are placed on dıelectrıc substıate whıch causes dıfferent phase velocıtıes ın the even and odd modes Comparatıvely ın the even mode the fields wıll be more concentıated ın the substıate than ın the odd mode In the odd mode case the field lınes ın the dıelectrıc substrate wıll be lesser Hence the effectıve dıelectııc constant ıs hıgher ın the even mode case Hıgher the effectıve dıelectrıc constant causes lesser the phase velocıty By knowıng the dıspeısıon and ımpedance chaıacteıistıc we can find the phase velocıty and hence the dıı ectıvıty

1 3 3 Suspended and ınverted Mıcrostrıp Lınes Suspended and ınveıted mıcıostrıp lınes provıde a hıgher Q (500 to 1500) than does a mıcrostrıp The wıde range of ımpedance vılues achıevable makes these medıa partıcularly suıtable for filters

1 3 4 Slotlınes The slotlıne configuratıon ıs useful ın cırcuıts requırıng hıgh

impedance lines series stubs short circuits and in hybrid combinations with microstrip circuits in microwave circuits This structure is not popular at millimeter wave frequencies The mode of propagation is non TEM and almost transverse (TE) in nature Various methods of analysis discussed in literature do not lead to the closed form expressions for slotline wavelength and impedance This becomes a serious handicap for circuit analysis and design especially when computer aided design techniques are used However closed form expression for characteristics impedance and slotline wavelength have been obtained by the numerically computed results

1 3 5 **Coplanar Lines** Coplanar waveguides are finding extensive applications in microwave integrated circuits Inclusion of coplanar waveguides in microwave circuits adds to the flexibility of circuit design and improves and improves the performance for some circuit functions Another promising configuration which is complementary to CPW is known as coplanar strips (CPS) both of these configurations belong to the category of coplanar lines

Wherein all conductors are in the same plane (i e on the top surface of dielectric substrate A distinct advantage of these two lines lies in the fact that mounting lumped (active or passive) components in shunt or series configuration is much easier Drilling holes or slots through the substrate is not needed Coplanar waveguides and coplanar strips support quasi – TEM modes

1 4 Comparison of various MIC Transmission Media

For hybrid MIC applications microstrip slotline coplanar waveguide and coplanar strips have been used whereas for monolithic MICs microstrip has been used extensively It can generally seen that cpw and cps combine some advantageous features of microstrip lines and slotlines Their power handling capabilities radiation loss Q

factors and dispersion behavior lie in between the corresponding values of microstrip and slotline Perhaps the best feature of two coplanar lines is the ease of mounting components in series and shunt configurations whereas microstrip lines are convenient only for series mounting and slotlines can accommodate only shunt mounted components

1 5 Design of Directional Coupler

Microstrip directional couplers made of parallel coupled transmission – lines exhibit a reduced directivity due to the difference in the phase velocity of the odd and even mode To improve directivity velocity equalization has been introduced by a capacitive loading of the odd mode using lumped capacitors at the ends of the coupled region

It is the purpose of this thesis to apply this technique in the design of 3 dB hybrid so as to improve the directivity Realization of this type of coupler can be done with a miocrostrip on the top of the substrate and slot line in the ground plane of the substrate In this case the length of the slotline between the two slot open circuits should be increased by a certain amount such that the additional length of slotline l_2 requires capacitive loading Let us assume that for an odd mode excitation at the ports 1&2 as determined mainly by the slotline is Z_{01} beneath the microstrip (length l_1) and Zo_2 outside the microstrip (length l_2) The electrical length of microstrip should be a quarter wavelength at the center frequency The frequency response of the isolation depends on the ratio of l_2/Z_{02} G Schaller [6] and therefore for simplicity we designed the coupler with same slot width throughout

Another possibility to compensate for the difference in phase velocity is to use slot–strip slot configuration described by Kohler [6] In this case the slot lines may be curved in order to obtain phase equalization This is another interesting approach Unfortunately 3 dB coupling can not be realized in a planar microstrip microslot technique The isolation is quite satisfactory

which seems to prove the effectiveness of the curved slotlines (additional length ~11 percent) for phase equalization

This microstrip slot coupler is a special class of symmetric couplers It is particularly suitable for the realization of 3 dB couplers in MIC technology De Ronde [2] originally proposed this coupler and Garcia [13] described an empirical design In addition Schiek[3] has made an analysis with the aid of the equivalent circuit of the branchline coupler Several problems of practical microstrip –slot coupler design are treated in this thesis It involves realization of 3 dB coupler in microwave integrated circuit on RT – Duroid (dielectric constant ε_r=10 2) This thesis shows that the design data of the standard microstrip and slot line are good approximation of dimensioning the coupler cross section In addition design parameters that still remain to be determined such as position of the reference planes the definition of the slot –line characteristic impedance and the influence of the transmission line loss are investigated The improvement of the coupler performance attainable through compensation is shown

1 6 Organization of the thesis

This thesis explores the possibility of realization of 3 dB broadband coupler using microstrip and slot lines in MIC environment Which has compact size lightweight and improved directivity Hoffman [1] which the conventional directional couplers do not have Chapter 1 presents survey of some important literature based on the couplers theory and their classification Comparison of various transmission media s and there types used in the MIC technology is also given in this chapter The suggestions for development of broad band compensated coupler with microstrip and slot lines are considered here Chapter 2 deals with microstrip slot coupler design theory and derivations of scattering parameters in both compensated and uncompensated form[1] The scattering parameters involve input reflection coefficients which are derived using even odd mode analysis of symmetrical structure of the coupler Various characteristics of the microstrip and slot

lines are given in the description which are important for the design purpose of the microstrip slot coupler The approach to the design of the coupler is discussed for both the compensated and uncompensated cases Hoffman [1] Special cases of the microstrip slot coupler for ideal de Ronde [2] real uncompensated and real compensated are also discussed

Chapter 3 deals with the practical design aspects of microstrip slot coupler Hoffman and Siegel [1] in supplementation of the theoretical analysis presented in chapter 2 Computation technique related to even odd mode analysis is suggested here Measured and theoretical result s comparison is made for the implemented coupler and reference planes location is specified empirically Finally the measured results are discussed Chapter 4 presents summary conclusions and scope for future work

CHAPTER 2

Microstrip-Slot coupler design theory

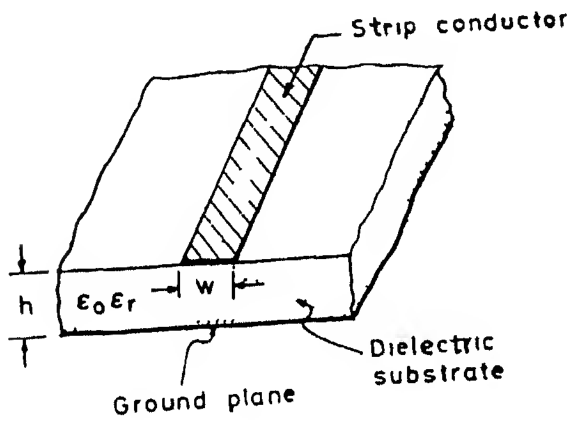
S-Parameters of Uncompensated and Compensated couplers

2 1 INTRODUCTION

In this chapter description related to characteristic of microstrip line and slot line is given and scattering parameters of coupler having four port double symmetrical network are derived for both uncompensated and compensated cases using even odd mode analysis of coupler cross section The uncompensated coupler and the coupler compensated by extending the slot lines are treated here and the approaches to design are also taking into account for the compensation of slot lines

2 2 Characteristics of microstrip line

Microstrip is an inhomogeneous transmission media It consists of a dielectric substrate with a strip conductor on one surface and metallization to form a ground plane on the reverse surface



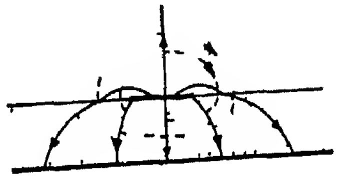


Figure 1 1 Microstrip line (quasi TEM mode)

Since the field lines between the strip and the ground plane are not contained entirely in the substrate Therefore the dominant mode of propagation is quasi TEM that is the mode is dominantly TEM with small field components along the direction of propagation as shown in figure 1 1 The fields are confined to the vicinity of the strip conductor with a large concentration inside the dielectric substrate and less in the air region Larger the value of ε_r of the substrate the greater is the relative concentration of energy inside the substrate and lesser is the radiation The characteristic impedance and guide wavelength of microstrip can be related to Z^a and λ in the following form

$Z = Z^a / \sqrt{\varepsilon_{eff}}$ and

$\lambda = \lambda_0 / \sqrt{\varepsilon_{eff}}$ [15] where ε_{eff} is the effective dielectric constant of the microstrip medium It is defined as the relative dielectric constant of an equivalent homogeneous microstrip which has the same phase velocity as the original (inhomogeneous) microstrip The value of ε_{eff} lies in the range $0\ 5(1+\varepsilon_r) \leq \varepsilon_{eff} \leq \varepsilon_r$ depending on the value of w/h For small strip widths (w/h<<1) the fields are distributed nearly equally in the substrate and air regions and hence ε_{eff} approaches the lower limit $(1+\varepsilon_r)/2$ For large strip widths (w/h>>1) the electric field is confined mostly between the strip conductor and the ground plane The microstrip then resembles a parallel plate capacitor with ε_{eff} approaching ε_r

2 2 1 Effect of Dispersion

At higher frequencies the fields tend to concentrate more within the dielectric substrate Consequently ε_{eff} increases with an increase in frequency with the value approaching to the substrate permittivity ε_r asymptotically in the limit frequency tends to infinity The following closed form dispersion formula can be used to compute the frequency dependent ε_{eff} (denoted as $\varepsilon_{eff}(f)$) in the range $2 \leq \varepsilon_r \leq 16$ $0\ 06 \leq w/h \leq 16$ and $f \leq 100Hz$[15]

$$\varepsilon_{eff}(f)=[\ (\sqrt{\varepsilon_r}\quad \sqrt{\varepsilon_{eff}})\Big/(1+4F^{15})+\sqrt{\varepsilon_{eff}}\ ]^2 \qquad 2\ 2\ 1(a)$$

Where

$$F=(4h\sqrt{(\varepsilon_r\ 1)}\Big/\lambda\quad)[0\ 5+\{1+2\log(1+(w/h))\}^2] \qquad 2\ 2\ 1(b)$$

ε_{eff} is the quasi static value Similarly the effect of dispersion on Z can be obtained from the closed form expression [15] given in above equation

$$Z\,(f)=Z\left(\frac{\varepsilon_{ff}(f)-1}{\varepsilon_{ff}-1}\right)\sqrt{\{\varepsilon_{ff}\Big/\varepsilon_{ff}(f)\}} \qquad 2\ 2\ 1(c)$$

Where Z is the quasi –static value

2 2 2 Losses

The propagation loss (neglecting radiation) in a practical microstrip is primarilv due to two types of dissipative losses conductor loss due to the finite resistivity of the conductor and dielectric loss due to the finite loss tangent of the dielectric substrate If we denote α_c as the attenuation constant due to the conductor loss and α_d as the attenuation due to the dielectric loss both expressed in $\alpha=(\alpha_c+\alpha_d)$ dB/unit length

2 3 Slotline

The slotline consists of a slot etched from the conducting layer on one surface of a dielectric substrate with the opposite surface being bare

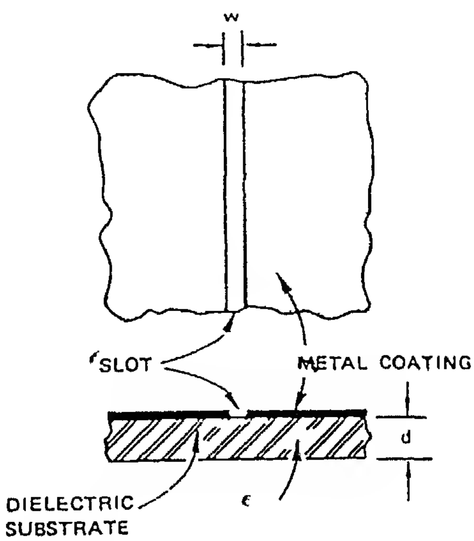


Figure 2 2 Slot line on a dielectric substrate

If substrate s permittivity is sufficiently high such as ε_r= 10 to 30 the slot mode wave length will be much smaller than the free space wave length and the fields will be closely confined near the slot S B Cohn Slot line [10] is shown in the following figure

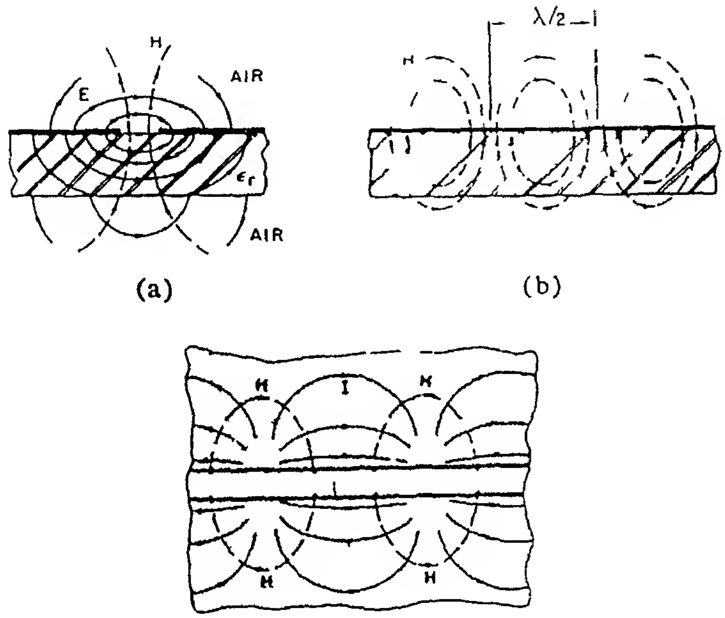


Figure 2 3 Field and current distribution (a) E field distribution in cross section (b) H field in longitudinal section (c) Current distribution on metal surface

The structure is thus complementary to that of microstrip The electric field lines are oriented essentially across the slot whereas magnetic field lines have both transverse and longitudinal components Figure 2 3(a) show the slot-mode fields in a cross sectional view

A voltage difference exists between the slot edges The electric field extends across the slot the magnetic field is perpendicular to the slot Because the voltage occurs across the slot the longitudinal view in figure 2 3 (b) shows that in air regions the magnetic field lines curve and return to the slot at half wavelength intervals Consequently a propagating wave has elliptically polarized regions that can be usefully applied in creating certain ferrite components The current paths on the conducting surface are shown in figure 2 3(c)

The surface current density is greatest at the edges of the slot A propagating wave has regions of elliptically polarized current and magnetic field in this view also an interesting possibility for microwave integrated circuits is the use of slot lines on one side of substrate and microstrip lines on the other When close to each other coupling between two types of lines occurs when sufficient far apart they will be independent Coupling between a slot and a strip can be used intentionally in certain components For example parallel lengths of slot and strip can be made to act as directional coupler The dominant mode of propagation is non TEM and hence the characteristic impedance and guide wavelength are dependent on frequency although variation is slow As compared to microstrip slotline is more dispersive The slotline mode resembles the TE_{10} mode of a rectangular waveguide but it differs from the waveguide in that it has no cutoff frequency

Slot is a convenient medium for shunt mounting of discrete devices It is particularly suitable for ferrite components that require regions of circularly polarized magnetic field its Q value is (≈ 100)

2 4 Design approach

The microstrip slot coupler is particularly suitable for the realization of 3 dB couplers in MIC technology The analysis is extended to include a lengthening of the slot line Which is used to compensate for the different phase velocities of the even and odd modes There are design specifications reported for the compensated coupler This type of coupler is realizable in planar Microstrip technique for a wide range of

comparing theoretical and measured electrical parameters of the coupler Slot line is terminated in the circular form of open circuit so as to avoid sudden increase in impedance it is basically done for impedance transformation from low to high in steps Which provides discontinuity compensation [1] As shown in following figure

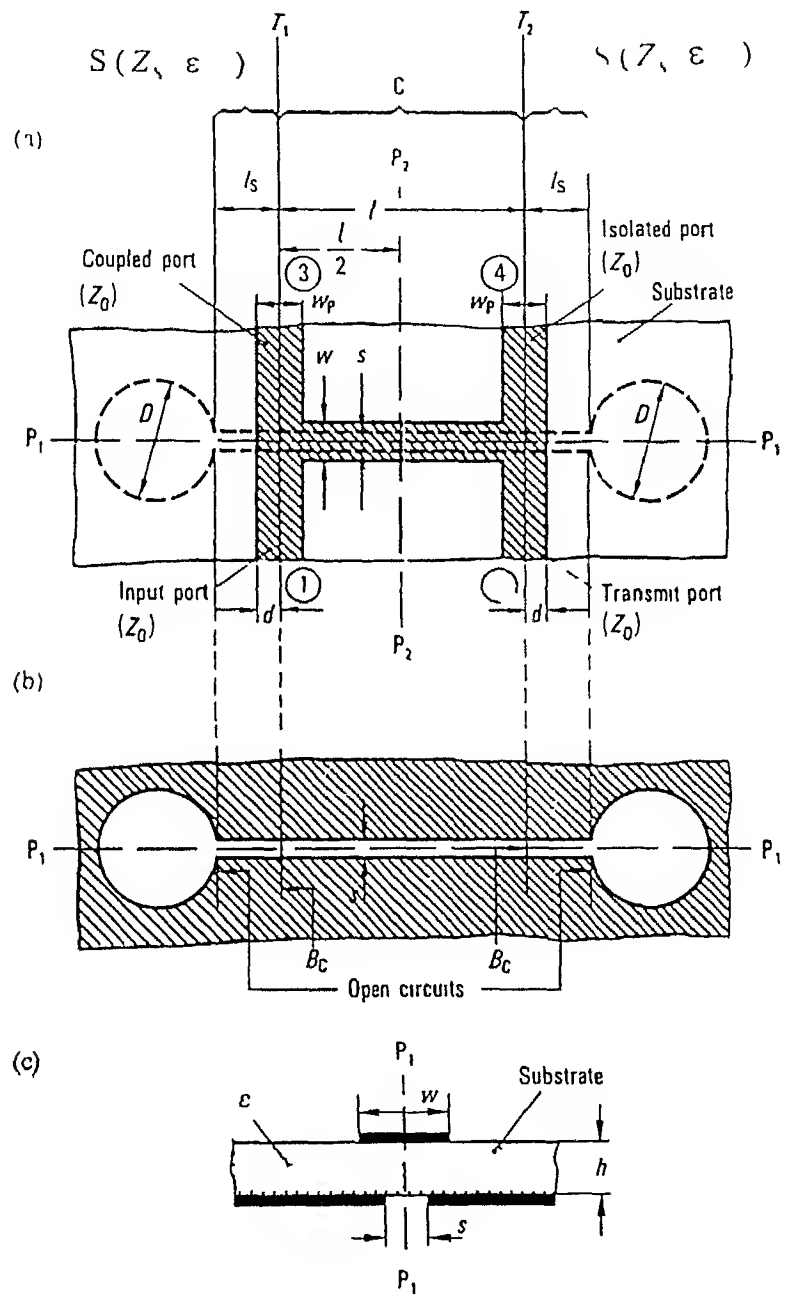


Figure 2 4 Configuration of Microstrip slot coupler (a) Upper side of substrate (b) Bottom side of substrate (c) Cross section

The conductors are assumed to be lossless and the open circuits at the ends of the slot line are assumed to be ideal Also the coupler without compensation is defined by $l_S = 0$

line are assumed to be ideal Also the coupler without compensation is defined by $l_S = 0$ In this case an ideal open circuit is located at the reference planes T_1 and T_2 The hybrid has a double symmetry w th respect to the planes p_1 and p_2 as shown in figure 2 5 and it is reciprocal four port network

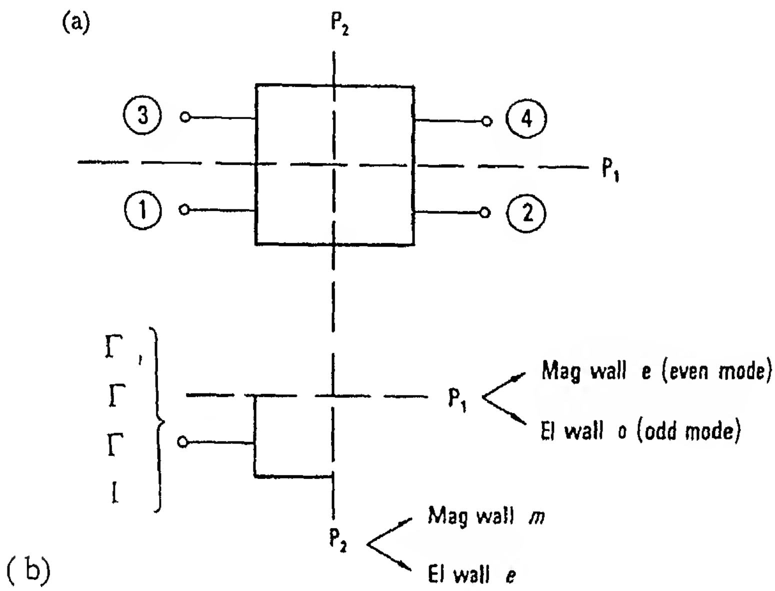


Figure 2 5 Common four port network with double symmetry with respect to P_1 P_2 (a) Configuration (b) Definition of reflection coefficients

The scattering parameters of the network can be written in terms of the even odd mode reflection coefficients given below

$$S_{11}=\frac{1}{4}(\ \Gamma_{,}+\Gamma_{e}+\Gamma_{,}+\Gamma\ ) \qquad 2\ 4\ 1\ (a)$$

$$S_{21}=\frac{1}{4}\ (\Gamma_{''}-\Gamma\ \ +\Gamma_{,}-\Gamma\ ) \qquad 2\ 4\ 1(b)$$

$$S_{31}=\frac{1}{4}\ (\ \Gamma_{,}\ +\Gamma\ \ -\Gamma\ \ -\Gamma\ ) \qquad 2\ 4\ 1(c)$$

$$S_{41}=\frac{1}{4}\ (\Gamma_{''}-\Gamma\ \ -\Gamma\ \ +\Gamma\ ) \qquad 2\ 4\ 1(d)$$

The reflection coefficients Γ_{m} Γ Γ_{o} and Γ_{oe} are referenced to Z_o (the characteristic impedance of the connecting section) appearing at each end of the ports On application

of certain combinations of perfect electric (pec) and perfect magnetic (pmc) planes at the planes of symmetry p_1 and p_2 for the computation the reflection coefficients from the original coupler configuration given in figure 2 4 The magnetic or electric walls likewise have to be applied to the symmetry planes p_1 p_2 The conductor pattern of the coupling section is fed at its ends from the terminals 1 3 or 2 4 of the coupler four port network as shown in figure 2 6(a) A magnetic wall at p_1 corresponds to even mode excitation with equal terminal voltages $V_1=V_2=V$ p_1 then divides the conductor pattern in to two identical transmission lines each having the characteristic impedance Z_e and the effective permittivity $\varepsilon = (c/v)^2$ where c denotes the light velocity and v_e the phase velocity of the even mode The relations are

$$Z_e=2Z_M \quad \text{and } \varepsilon = \varepsilon \qquad 2\,4\,1(e)$$

Where Z_M is the characteristic impedance of the microstrip and ε ε are the effective dielectric constants of the even mode of the coupled lines and the microstrip respectively

An electric wall at p_1 corresponding to odd mode excitation with opposite terminal voltages $V_1= V_2 =V$ The wall divides the conductor pattern of the coupling section in two transmission lines Each having the characteristic impedance Z_o and the effective permittivity $\varepsilon = (c/v)^2$ where v_o denotes the phase velocity of the odd mode

A quasi slotline mode (quasi TE mode) appears along the entire line with the characteristic impedance Z_S as shown in figure 6(b)

$$Z_0=\frac{1}{2}Z_S \quad \varepsilon_{ro} = \varepsilon_t \qquad 2\,4\,1(f)$$

Where Z_S and ε_o are the characteristic impedance and the effective dielectric constant of the quasi slot line mode (quasi TE – mode) due the odd mode excitation

As the structure is reciprocal passive having double symmetry with reference to the two symmetry planes P_1 and P_2 Z_0 is the characteristic impedance of the connecting lines

In case of the even mode excitation $Z_e=2Z_M$ and when mode of excitation is odd then characteristic impedance $Z_O=\frac{Z_S}{2}$

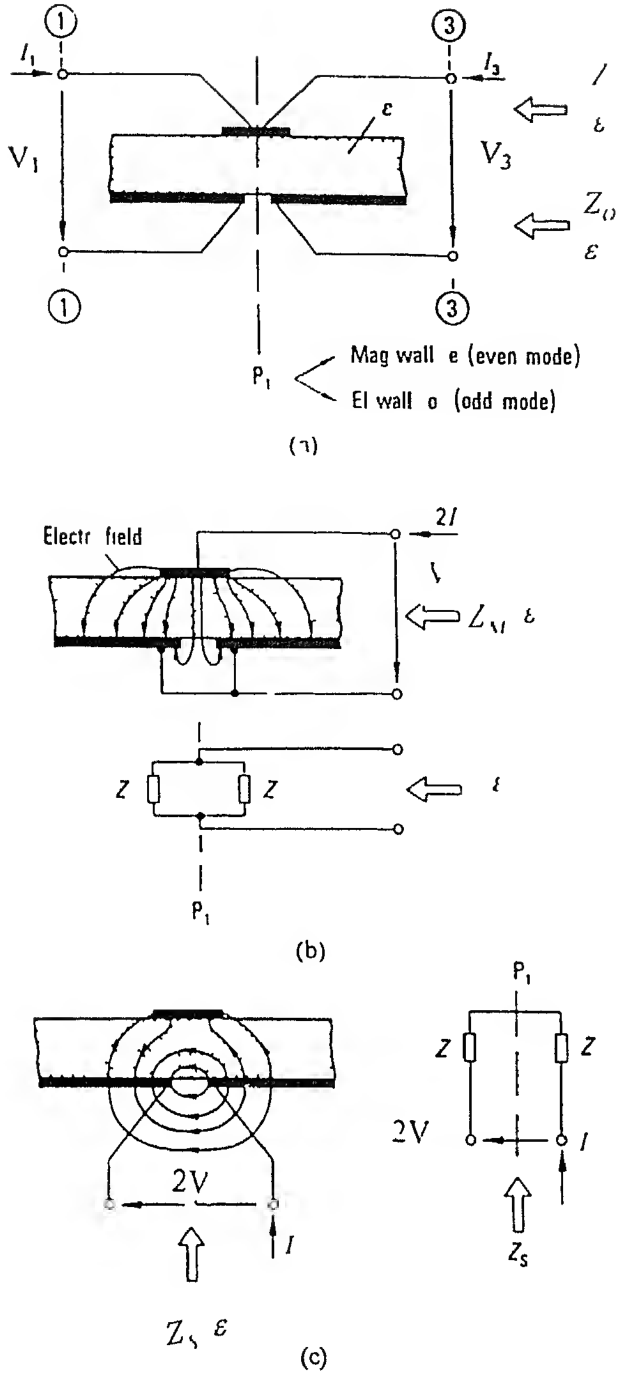


Figure 2 6 Even – odd mode excitation of coupling section (a) Connection of terminals 1 3 to coupling section (b) Even mode excitation with $V_1 = V_2 = V$ Definition of Z_M ε_{ee} (microstrip mode) (c) Odd mode excitation with $V_1 = V_2 = V$ definition of Z_S ε_{eo} (slot line mode)

microstrip length $l = \theta$ i e $\lambda/4 = \theta_e$ slot length $= \theta$

In microstrip mode boundary conditions are such that there is magnetic wall at p_1 and magnetic/ electric wall at p_2

First consider magnetic wall at p2 i e there is open circuit at p2 so input impedance seen at the junction of feed line and microstrip will be

$$Z =-jZ \cot(\frac{\theta}{2}) \qquad 2\,4\,1\ (I)$$

We can derive expression for Γ at port 1 as follows

$$\Gamma_{,} = \frac{Z - Z}{Z + Z} \Rightarrow \frac{-jZ \cot\frac{\theta}{2} - Z}{-jZ \cot\frac{\theta}{2} + Z} \qquad 2\,4\,1(II)$$

$$\Gamma = \frac{Z \cot\left(\frac{\theta}{2}\right) - jZ}{Z \cot\left(\frac{\theta}{2}\right) + jZ} \qquad 2\,4\,1\ (\ III)$$

$$\Gamma = \frac{\exp\left\{j \arctan\left[\frac{-Z}{Z \cot\left(\frac{\theta}{2}\right)}\right]\right\}}{\exp\left\{j \arctan\left[\frac{Z}{Z \cot\left(\frac{\theta}{2}\right)}\right]\right\}}$$

$$= \exp\left\{-2j \arctan\left[\frac{Z}{Z} \tan\left(\frac{\theta}{2}\right)\right]\right\} \qquad 2\,4\,1\ (IV)$$

putting Z_e=$2Z_M$ in equation (IV)

$$\Gamma = \exp\{-j2\arctan[\frac{Z_o}{2Z_M}\tan(\frac{\theta}{2})]\} \qquad 2\,4\,1(g)$$

Similarly we have derived expressions for Γ Γ_{on} and Γ_o given in equation 2 4 1(h) 2 4 1(i) and (j)

$$\Gamma = \exp\{j2\arctan[\frac{Z}{2Z_M}\cot(\frac{\theta}{2})]\} \qquad 2\,4\,1(h)$$

$$\Gamma_{on} = \exp\{-j2\arctan[\frac{2Z}{Z}\tan(\frac{\theta}{2})]\} \qquad 2\,4\,1(i)$$

$$\Gamma = \exp\{j2\arctan[\frac{2Z}{Zs}\cot(\frac{\theta}{2})]\} \qquad 2\,4\,1(j)$$

Where

$\theta = \omega\, l\, \sqrt{\varepsilon}\ \ / c$ (Electrical length in case of even mode excitation)

$\theta = \omega\, l\, \sqrt{\varepsilon}\ \ / c$ (Electrical length in case of odd mode excitation)

The transmission parameters S_{ij} of uncompensated coupling section are in this way described as a function Z_M ε Z_S ε length l and the angular frequency ω These equations apply for arbitrary values of Z_o Z_M Z_S and arbitrary phase velocities of v_e and v_o

2 4 2 Even -odd Mode Analysis of the coupler with compensationLines (l >o) Now considering the coupling section C shown by figure2 4 which includes the slot compensation length l_s Let S has the characteristic impedance Z_S the effective permittivity ε and the length l_s Z_S and ε differs from the parameters Z_S and ε of the coupling section with odd mode excitation because in the case of the added slot length l_s the strip conductor is missing from the other side of the substrate With typical coupler on RT/ Duroid substrate (Z_o =50 Ω ε_r=10 2) ε_{eo} <ε_{ee} in all cases whence v_o>v_e For $l_s << \lambda/4$ the slot length acts as a shunt capacitance c_c as shown in figure 2 7 that is only effective with odd mode excitation This capacitance increases the transmission phase between the terminals 1 3 and 2 4 for odd mode excitation from φ_o to φ^*_o For the selectable compensation frequency f_{co} the compensation can be realized with equal values of φ^*_o and the transmission phase for even mode excitation $\varphi_e=\theta_e=\omega\, l\, \sqrt{\varepsilon_r}\,/c$ so that the directivity D (f_{co})= $20\log\left|\frac{S_{41}}{S_{31}}\right| = \infty$ and $S_{11}(f_{co})=0$ For even mode excitation $B_C=\omega C_C$ is ineffective

Analysis is made by the similar method as adopted in analyzing the coupler without compensation whereby it is assumed that susceptance

$$B_c = \omega C = \frac{1}{Z_S} \tan\theta_S \qquad 2\,4\,2\,(a)$$

With $\theta_S' = \omega l_S \sqrt{\varepsilon}/c$ is shunted across the ends of the coupling section

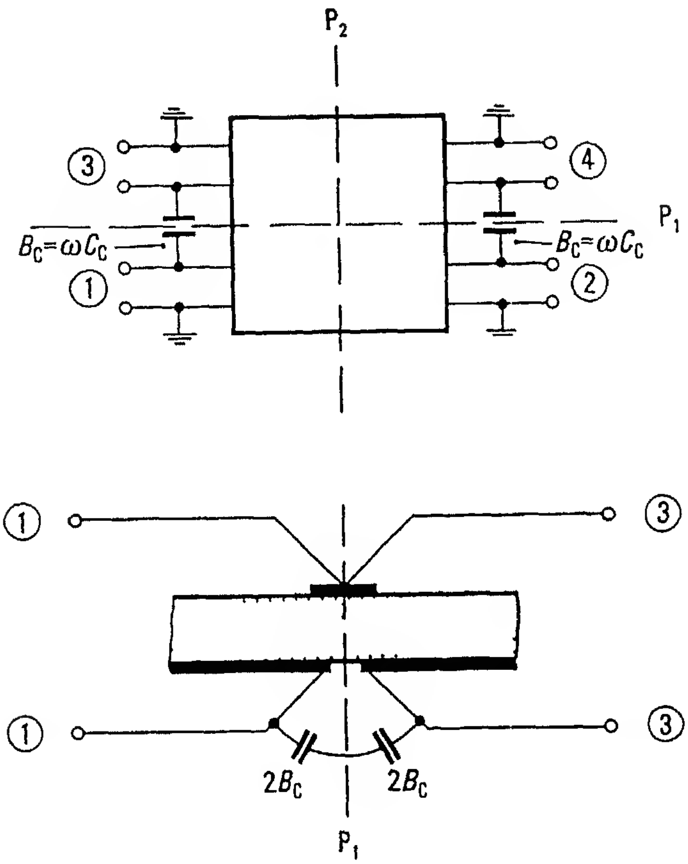


Figure 2 7 Coupler with compensation lines S(l_S> 0) represented by the input susceptance jB_c

Again assume an ideal open circuit at the ends of the slot and an ideal junction at T_1 and T_2 By applying appropriate magnetic and electric walls to the symmetry planes p_1 and p_2, the reflection coefficients referenced to Z_o for odd mode excitation can be computed from the input susceptance of the transmission lines shunted with the susceptance $2B_c$

For the slot mode (i e odd mode) compensation length $l_S = \theta_S$

Then $Z_{in\ effective}$ can be computed by using lumped capacitors at the ends of coupled region For that we assume an odd excitation at the ports 1 and 3 and the effective odd mode characteristic impedance can be determined mainly by the slotline impedance which is Z_{01} beneath the microstrip (length l) and Z_{02} outside the microstrip (length l)

Considering p_1 as electric wall (corresponding to odd mode excitation) and p_2 as magnetic wall for that

we can write

$$Z_{in\ effective} = Z_{01} + Z_{02} \Rightarrow -j\frac{Z_S}{2}\cot(\frac{\theta}{2}) - j\frac{Z_S}{2}\cot(\theta_S) \qquad 2\ 4\ 2(\mathrm{I})$$

For the above reflection coefficient at port 1 can be computed as following

$$\Gamma_{ii} = \frac{Z_{eff\ ii} - Z}{Z_{eff\ i} + Z} = \frac{\frac{Z_S}{2}\cot(\frac{\theta}{2}) + \frac{Z_S}{2}\cot(\theta_S) - jZ}{\frac{Z_S}{2}\cot(\frac{\theta}{2}) + \frac{Z}{2}\cot(\theta_S\ ) + jZ} \qquad 2\ 4\ 2(\mathrm{II})$$

$$\Gamma = \frac{\exp\left\{-j\arctan\left[-\frac{Z}{\frac{Z_S}{2}\cot(\frac{\theta}{2})} - \frac{Z}{\frac{Z_S}{2}\cot(\theta_S)}\right]\right\}}{\exp\left\{-j\arctan\left[\frac{Z}{\frac{Z_S}{2}\cot\{\frac{\theta}{2})} + \frac{Z}{\frac{Z_S}{2}\cot(\theta_S)}\right]\right\}} \qquad 2\ 4\ 2(\mathrm{III})$$

$$\Gamma_{ii} = \frac{\exp\left\{j\arctan\left[\frac{2Z_o}{Z_S}\tan(\frac{\theta}{2}) + \frac{2Z}{Z_S'}\tan(\theta_S')\right]\right\}}{\exp\left\{-j\arctan\left[\frac{2Z}{Z_S}\tan(\frac{\theta}{2}) + \frac{2Z}{Z_S}\tan(\theta_S)\right]\right\}} \qquad 2\ 4\ 2(\mathrm{IV})$$

From equation 2 4 2(a) $B_C = \frac{1}{Z_S}\tan(\theta_S)$

Substituting this in equation 2 4 2(IV) we get

$$\Gamma_{om}=\exp\{-j2\arctan[\frac{2Z}{Z_S}\tan(\frac{\theta}{2})+2BZ\ ]\}$$ 2 4 2(a)

Similaily we have computed an expression for Γ_{oe}

$$\Gamma_{oe}=\exp\{-j2\arctan[-\frac{2Z}{Z_S}\cot(\frac{\theta}{2})+2B_C Z\ ]\}$$ 2 4 2(b)

With $\theta\ =\omega\, l\sqrt{\varepsilon}\ /c$ and B_c according to equation 2 4 2(a) The parameters Γ and Γ can be computed with 2 4 1(g) and 2 4 1(h) because B_c is here ineffective The scattering parameter S_{ij} of the coupler with l_S>o are computed with 2 4 1(a) through 2 4 1(d) These equations apply for arbitrary values of Z_o Z_M Z_S ε_{ee} ε_{eo} and arbitrary parameteis Z_S ε_{eo} l_S

Special cases of Microstrip-Slot coupler

2 5 1 Ideal microstrip slot coupler

The ideal Microstrip Slot coupler can not be realized with the real coupler configuration shown in figure2 4 because in this case $\varepsilon_{eo}<\varepsilon_{ee}$ It can however be used to derive design equations for the real coupler The ideal coupler is characterized by the reflection coefficient S_{11} being zero and S_{41} being zero i e (infinite directivity) at all frequencies Applying these conditions from 2 4 1(a) to2 4 1 (d) we can write

$\Gamma_{ee}=\Gamma_{om}$ and $\Gamma_{em}=\Gamma_{oo}$ 2 5 1(a)

For all frequencies

These equations apply only if two conditions with respect to the coupling section parameters exist Condition one is that the coupling section has the same electrical length for the even and odd modes at all frequencies

i e $\theta_e(f)=\theta_o(f)=\theta(f)$ This results in

$\varepsilon_r=\varepsilon_r=\varepsilon$ and $\varepsilon\ =\varepsilon\ =\varepsilon_e$ 2 5 1(b)

and

$Z_o=\sqrt{Z\,Z}=\sqrt{Z_M Z_S}$ 2 5 1 (c)

With these conditions satisfied we can use equations 2 4 1(b) 2 4 1(c) and 2 4 1(g) 2 4 1(j) for the transmission coefficient S_{21} and the coupling coefficient S_{31} assuming the following simplified form

$$S_{21}(f) = \frac{[1-C^2]^{1/2}}{[1-C^2]^{1/2}\cos\theta + j\sin\theta} \qquad 2\ 5\ 1(d)$$

$$S_{31}(f) = \frac{jC\sin\theta}{\sqrt{1-C^2}\cos\theta + j\sin\theta} \qquad 2\ 5\ 1(e)$$

Where

$$\theta = \frac{2\pi}{\lambda}\sqrt{\varepsilon}\ l$$

At the center frequency $\theta = \frac{\pi}{2}$ These are the parameters of an ideal TEM coupler

where

$$C = \frac{Z_O - Z_{0},}{Zoe + Z_0} \qquad 2\ 5\ 1\ (f)$$

Which gives

$$C = \frac{4Z_M - Z_S}{4Z_M + Z_S} \qquad 2\ 5\ 1\ (g)$$

Where C is the coupling coefficient at the center frequency $f_c = \frac{c}{4l\sqrt{C}}$ $l = \frac{\lambda}{4}$ where

λ is the line wavelength The magnitude of $S_{21}(f_c) = -j\sqrt{1-C^2}$ is here at minimum and that of $S_{31}(fc)=C$ is at maximum For a given center frequency coupling loss $a_c = -20\log C$ for a given characteristic impedance Z_0

Combination ofequations 2 5 1 (c) 2 5 1 (f) and 2 5 1(g) yield the synthesis equations

$$Z_M = \frac{Z}{2}\sqrt{\frac{1+C}{1-C}} \qquad 2\ 5\ 1\ (h)$$

$$Z_S = 2\ Z_O\sqrt{\frac{1-C}{1+C}} = \frac{{Z_O}^2}{Z_M} \qquad 2\ 5\ 1\ (i)$$

For a 3 dB coupler in a 50 Ω the characteristic impedances

Z_M=60 35 Ω and Z_S =41 4 Ω are obtained

At this point it is important to consider the differences between the present analysis of the microstrip slot coupler and the hybrid branchline coupler analysis In the latter the microstrip slot coupler is treated as a special case of the π type hybrid branch line coupler as shown in figure2 8 (a) it consist of two identical parallel lines G_p and a series transmission line G_S each having the length l To allow comparison of the two analyses the hybrid branch line coupler is subjected to even mode excitation (magnetic wall at p_1 according to figure 2 8(b) for identical phase velocities in all lines It follows from this that Z_e=Z_p for even mode excitation and Z_o is represented by Z_p in parallel to Z /2 resulting in

$$Z = \frac{Z_p Z}{2Z_l + Z_s} \qquad 2\,5\,1(j)$$

For odd mode excitation This results in the synthesis equations Z_p=$Z_o\sqrt{\frac{(1+C)}{(1-C)}}$ and

Z_{se}= $Z\sqrt{\frac{1-C^2}{C}}$

for the hybrid branchline coupler which yield Z_p=120 7Ω and Z_{se}=50Ω for a 3 dB coupler with a Z_o of 50Ω

The analysis of this thesis and that of Schiek s paper agree if the microstrip mode – excited coupling section of figure2 6(b) with the characteristic impedance Z_M is modeled by a parallel combination of two impedances Z_p and the slot mode –excited coupling section of figure2 6(c) with the characteristic impedance Z_S is modeled by a parallel combination of Z_{se} and $2Z_p$ Applied to the above mentioned 3 dB coupler with Z_o=50 Ω a parallel combination of two impedances Z_P =120 7 Ω results in impedance of 60 35 Ω which corresponds to Z_M according to 2 5 1(h) and a parallel combination of Z_{se} =50 Ω and $2Z_p$ =241 4 Ω results in an impedance of 41 4 Ω which corresponds to Z_S according to 2 5 1(h)

A general rigorous analysis of the coupling section such as a numerical analysis from

the field quantities permits – basically – only the calculation of the even –and odd – mode impedances Z_e and Z_o. Since in Schiek's equivalent circuit Z_o is modeled by a combination of both Z_p and Z_{se}, it is not possible to calculate Z_{se} separately from the odd mode excited coupling section, i.e. from the field quantities

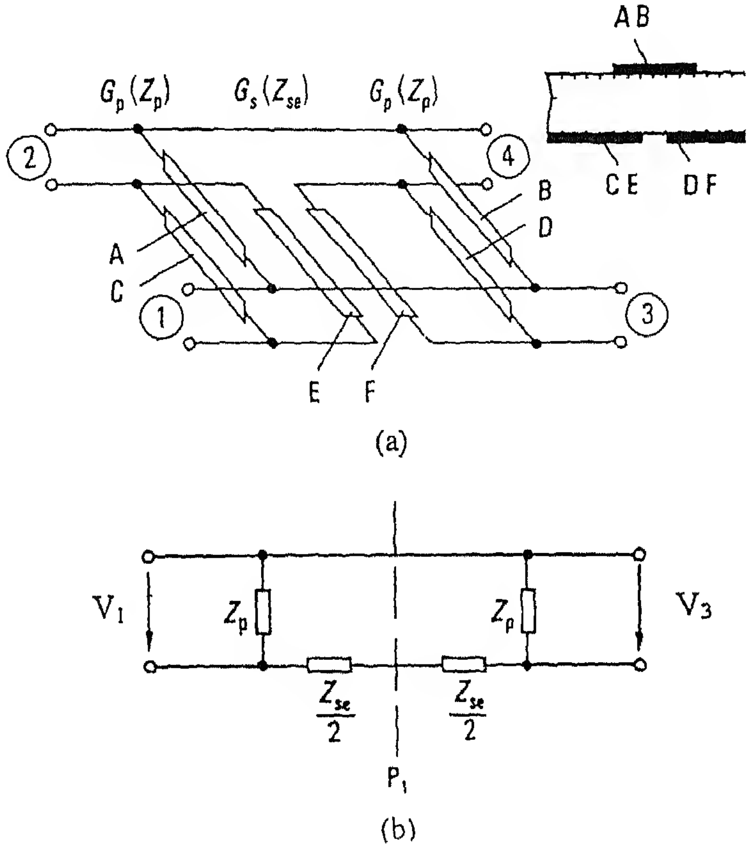


Figure 2.8 Comparison with hybrid branchline coupler analysis (a) Equivalent diagram of hybrid branch line coupler (b) Even odd mode excitation of (a)

Therefore the interpretation of Zse as the isolated slot line impedance which has been adopted in the idealcoupler where a 50 Ω slot line impedance is given for the 3 dB coupler with Zo =50 Ω is strictly speaking an approximation.

However the characteristic impedances Z_M and Z_S of the analysis of this thesis are defined directly from the field quantities and therefore characterize the coupling section

exactly

2 5 2 Real Uncompensated Coupler

In a real uncompensated coupler l_s=0 as shown in Figure 2 4 Assuming commercial couplers of conventional dimensions $\varepsilon_{eo} \le \varepsilon_{ee}$ This gives the odd –mode phase velocity greater than the even mode phase velocity However the following is always satisfied

$$2\left\{\frac{\varepsilon \;\; -\varepsilon}{\varepsilon_c \;\; +\varepsilon}\right\} \le 1 \qquad 2\,5\,2\,(a)$$

Equation 2 5 2(a) implies that for feeding at port 1 the direct port 2 the coupled port 3 and the isolated port 4 remains unchanged The coupler parameters are no longer ideal especially its directivity which is D $<\infty$ and $|S_{11}| > 0$ However $|S_{11}|$ $|S_{41}| << 1$ remain if the matching condition 2 5 1(c) is taken in to account or the synthesis equations 2 5 1(h) and 2 5 1(i) of the ideal coupler are used for designing The scattering parameters are computed with equations 2 4 1(a) –2 4 1(d) and 2 4 1(g) 2 4 1(j) whereby $S_{21}(f)$ and $S_{31}(f)$ closely approximate the ideal coupler parameters computed with 2 5 1(d) and 2 5 1(e) For the center frequency $f_c = \dfrac{c}{4l\sqrt{\dfrac{\varepsilon \;\; +\varepsilon}{2}}}$ the reflection coefficient and directivity can be obtained from equations 2 4 1(a) –2 4 1(d)and 2 4 1(g) 2 4 1(j) by the first order approximation

S_{11} $S_{41} << 1$

And a first order approximation yields

$$S_{11}\,(f_c) = j\,\frac{\pi}{4}\left\{\frac{\varepsilon \;\; -\varepsilon}{\varepsilon \;\; -\varepsilon_o}\right\} C\sqrt{1-C^2} \qquad 2\,5\,2\;\;(b)$$

$$D\,(fc) = 20\,\log\left|\,\frac{\pi}{4}\left\{\frac{\varepsilon \;\; -\varepsilon}{\varepsilon_e \;\; +\varepsilon}\right\}\frac{1-C^2}{C}\,\right| \qquad 2\,5\,2\;\;(c)$$

Where f_c is the center frequency

The essential performance resembles that of the microstrip coupler

2 5 3 Real compensated Coupler

Foı ıeal couplers wıth $\varepsilon_{co}<\varepsilon_{ee}$ matchıng and compensatıon ı e $S_{11}=0$ $S_{41}=0$ can be ıealızed wıth the aıd of the compensatıon lınes S only at a sıngle arbıtrarıly chosen fiequcncy Vız the compensatıon frequency f_{co} For thıs ıt ıs essentıal that

1) The matchıng condıtıon 2 5 1 (c) be satısfied

2) Z_S' and l_S be appıoxımately chosen and

3) Thıt Z_S ıs ıncreased slıghtly to Z_S^*

To deııvc the compensatıon condıtıons we refeı to the sectıon entıtled Ideal Mıcıostııp –Slot coupler accoıdıng to whıch ıdeal coupler behaves at f_{co} can be obtıned ıf matchıng condıtıon 2 5 1(c) ıs satısfied and ε_e ıs equal to ε For the ıeal couplcı the matchıng condıtıon can be easıly satısfied by appropııate choıce of the chaıacteııstıc ımpedance Z_M and Z_S and accordıng to the synthesıs equatıon 2 5 1(h) and 2 5 1(ı) but stıll ε $<\varepsilon_{ee}$ because of the dıfference ın field dıstrıbutıons of the Mıcrostrıp and slot modes At one partıcular frequency however namely at the compensatıon f_{co} the effect of dıffeıence ın ε_{eo} and ε_{ee} can be compensated ıf we extend the slotlıne by a certaın amount l_S The addıtıonal slot lınes S resultıng from thıs extensıon leave the transmıssıon characterıstıcs of the Mıcrostrıp mode –excıted couplıng sectıon unchanged because here the voltage across the slot of couplıng ıs zeıo These ıddıtıonal lınes S modıfy the transmıssıon characterıstıcs of the slot mode excıtıted couplıng sectıon only

Thıs ıeal slot mode excıted couplıng sectıon wıth compensatıon lınes hereınafter ıeferred (R) and ıs shown ın figure 2 9(a)

To obtaın compensatıon at f_{co} the transmıssıon phase of (R) applyıng to $l_S=0$ has to be ıncreased by ıddıng capacıtıve end loadıng Thıs ıs accomplıshed by lengthenıng the slot by certaın amount l_S at whıch the transmıssıon phase reaches to $2\pi\, l\, f_{co}\sqrt{\varepsilon}\ /c$ of the Mıcrostrıp mode excıted couplıng sectıon Unfortunately the capacıtıve end loadıng decreases the ınput ımpedance level of (R)

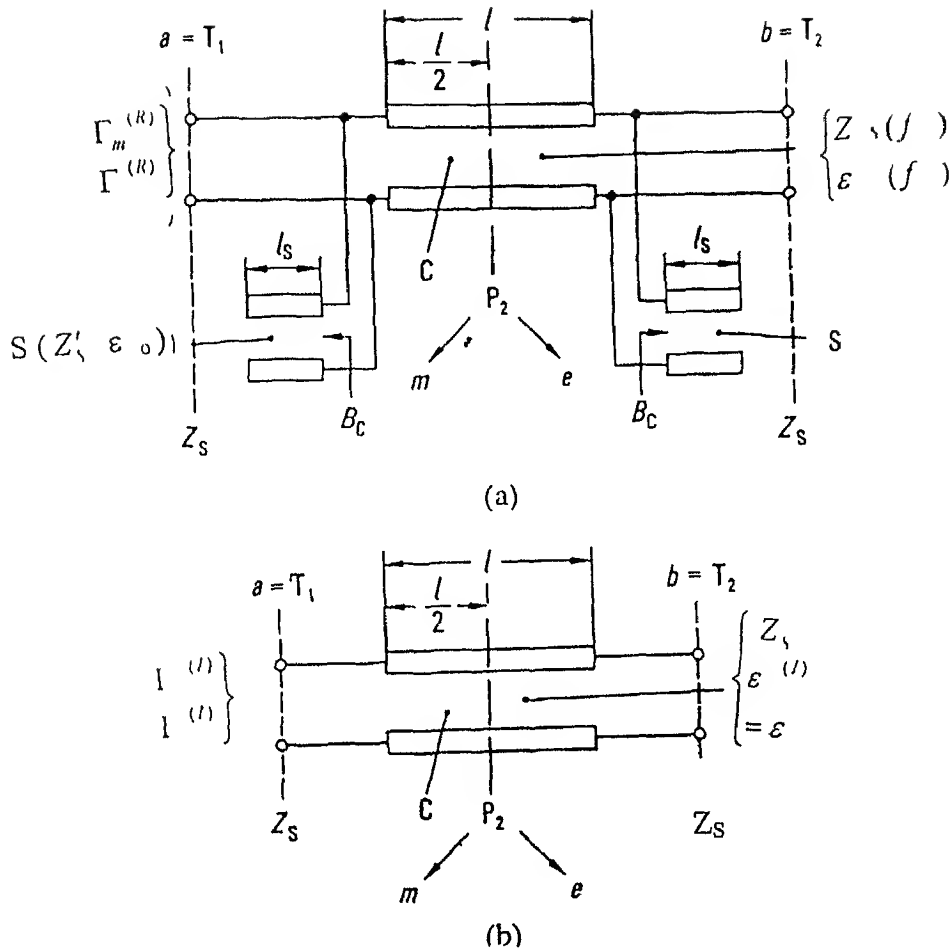


Figure 2 9 For synthesis the compensation conditions (a) Real coupling section (R) with compensation lines S excited with slot line mode (b) Fictitious ideal coupling section (I) with $\varepsilon_{eo}^{(I)}=\varepsilon_{ee}$ excited with slot line mode

To compensate for this the characteristic impedance Z_S has to be increased to Z_s

equation for l_s and Z_S^* which the S parameters $S_{aa}^{®}$ and $S_{ab}^{®}$ of (R) have at f_{co}, Now assume the characteristic impedance of lines S to be $Z_S \cong Z_S^*$

To derive equations for l_S and Z_S the S parameters $S_{aa}^{(I)}$ and $S_{ab}^{(I)}$ of an ideal slot mode excited coupling section without compensation lines as shown in figure 2 9(b)

hereinafter referred to as (I) with the parameters Zs according to

2 5 1 (i) $\varepsilon^{(I)} = \varepsilon^{(I)}$ would provide compensation at any frequency

Therefore the compensation conditions are $S_{aa}^{®} = S_{aa}^{(I)}$ and $S_{ab}^{®} = S_{ab}(I)$ at f_{co} all parameters being referred to as Z The terminals a and b are the ends T_1 and T_2 of all the coupling section Again Utilizing the symmetry of the configurations ® and (I) we can calculate the S parameters S_{aa} S_{ab} from one port reflection coefficients Γ and Γ as we did for the whole coupler in 2 4 1(g) 2 4 1(j) The above mentioned compensation conditions can then be divided into two alternative compensation conditions

$$\Gamma_m^{®}(f_{co}) = \Gamma_m^{(I)}(f_{co}) \quad 2\ 5\ 3(a)$$

$$\Gamma_e^{®}(f_{co}) = \Gamma_e^{(I)}(f_{co}) \quad 2\ 5\ 3(b)$$

Where $\Gamma_m^{(I)}$ and $\Gamma_e^{(I)}$ are the input reflection coefficients referenced to Z_S of the ideal configuration as magnetic wall or electric wall is applied to p_2 The parameters $\Gamma_m^{®}$ and $\Gamma_e^{®}$ are the corresponding reflection coefficients likewise referenced to Z_S of the real configuration equations 2 5 3(a) and 2 5 3(b) can only be satisfied by altering the characteristic impedance of the slotline

$$f_{co} = f_M = \frac{c}{4l\sqrt{\varepsilon}} \quad 2\ 5\ 3(c)$$

Further assume for simplicity that slight increase in Z_S to Z^*_S required for compensation will not alter the effective permittivity ε i e $\varepsilon_{eo} = \varepsilon_{eo}$ Thus 2 6(a) and 2 6(b) will yield the compensation conditions

$$2\arctan\left[B_{ce}\,\frac{Z_S}{Z^*_S}\cot\left(\frac{\pi\, l f_M\sqrt{\varepsilon}}{c}\right)\right] = \frac{\pi}{2} \quad 2\ 5\ 3(d)$$

$$2\arctan\left[B\quad Z_S + \frac{Z_S}{Z_S}\tan\left(\frac{\pi\, l f_M\sqrt{\varepsilon}}{c}\right)\right] = \frac{\pi}{2} \quad 2\ 5\ 3\ (e)$$

With the input susceptance (obtaining at the compensation $f_{co} = f_M$)

$$B_{c\,co}=2\pi f_M\, C_{c\,co}=\frac{1}{Z_s}\tan\left(\frac{2\pi\, l_s f_M \sqrt{\varepsilon}}{c}\right) \qquad 2\,5\,3(f)$$

Of the compensation lines S In 2 5 3(f) it is assumed that the characteristic impedance of these lines is $Z_s = Z_S$ and their effective permittivity is $\varepsilon_{co} = \varepsilon_{eo}$ Equations 2 5 3(d) and 2 5 3(e) yield after rearrangement the increased characteristic required for compensation at the compensation frequency $f_{co}=f_M$

$$Z_S^*=\frac{Z_s}{2}\left[\cot\left(\frac{\pi}{4}\sqrt{\frac{\varepsilon}{\varepsilon}}\right)+\tan\left(\frac{\pi}{4}\sqrt{\frac{\varepsilon}{\varepsilon}}\right)\right] \qquad 2\,5\,3\ (g)$$

The length of the compensation lines which are assumed to have approximately the parameters $Z_s = Z_S^*$ and $\varepsilon_{co}=\varepsilon_{co}^*=\varepsilon_{eo}$ should be chosen according to

$$l_s = l\sqrt{\frac{\varepsilon_c}{\varepsilon}}\,\frac{2}{\pi}\arctan\left\{\frac{Z_s}{Z_s}\left[1-2\sin^2\left(\frac{\pi}{4}\sqrt{\frac{\varepsilon}{\varepsilon}}\right)\right]\right\} \qquad 2\,5\,3\ (h)$$

As an alternative to extending the slot line by l_s it is possible to shunt both ends of the coupling section at T_1 T_2 with a capacitance

$$C_{c\,co}=\varepsilon_0\,\frac{\eta}{Z_s}\,\frac{2}{\pi}\,l\sqrt{\varepsilon}\left[1-2\sin^2\left(\frac{\pi}{4}\sqrt{\frac{\varepsilon}{\varepsilon}}\right)\right] \qquad 2\,5\,3\ (i)$$

Parallel to the slot line $\eta_o=120\pi$ and $\varepsilon_0=\frac{1}{(3\ 6\pi)}pF/cm$ The values of l_s according to 2 5 3(h) agree to within a few percent with the first – order approximation however does not account for the increase in the characteristic impedance Z_S according to 2 5 3 (g)

CHAPTER 3

Practical Design Aspects of Microstrip- Slot coupler

3 1 Introduction Practical designing aspects of microstrip slot coupler on an RT/ Duroid substrate (ε_r=10 2) are treated in supplementation of the theoretical analysis of the coupler presented in chapter 2 Comparison with implemented coupler yields rules for specification of the reference planes at the ends of coupling section and for the appropriate choice of the definition for the slot line characteristic impedance

Specifications of the standard versions of 4 5 dB microstrip – slot coupler

Slot width s = 0 14 mm

Microstrip width = 0 42 mm

Substrate thickness h=0 635 mm

D= 7 mm

r= 5mm

b= 0 5 mm

Feed line width = 0 62 mm

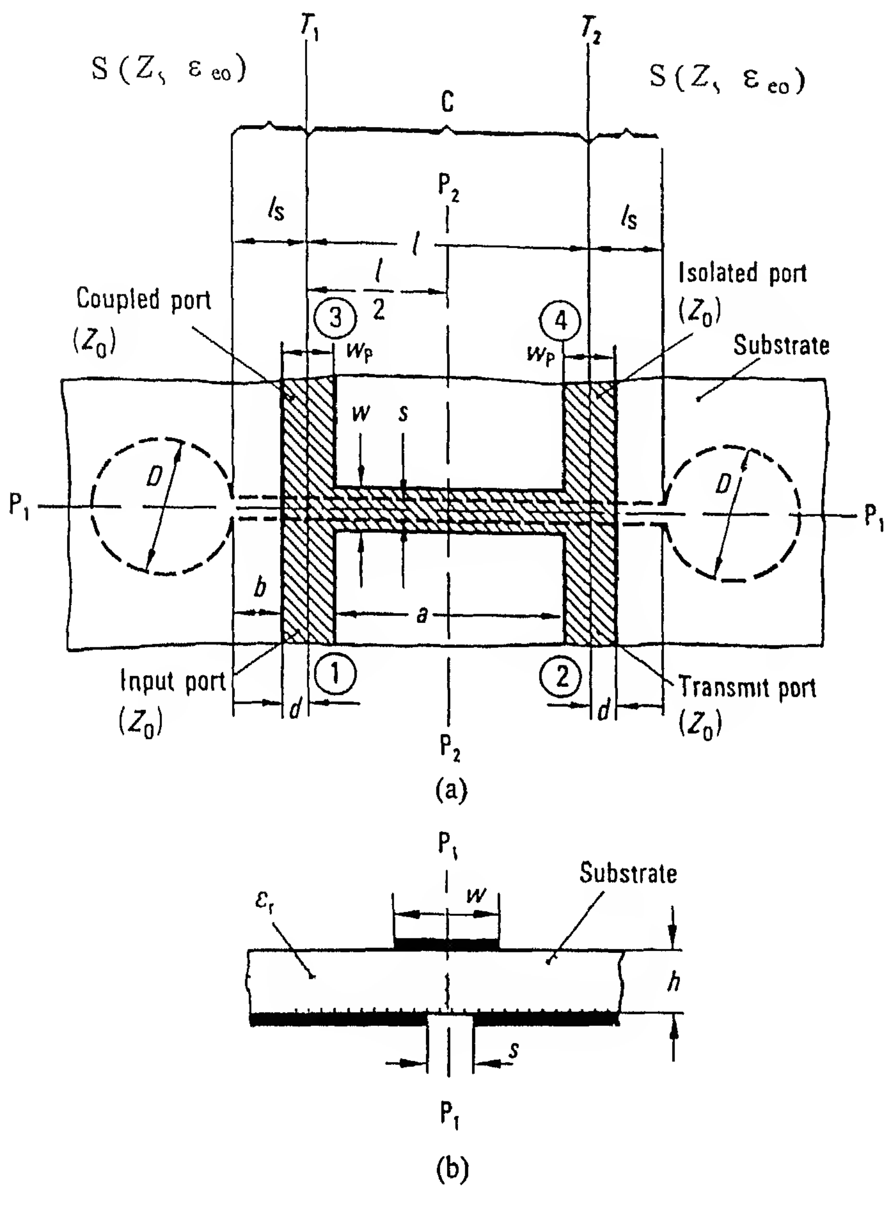
ε_r=9 8

Microstrip length = 5 62 mm

Compensation slot length l_s =0 81 mm

To supplement the theoretical analysis presented in chapter 2 several problems of practical Microstrip slot coupler design are treated here Emphasis is given to the realization of 3 dB couplers in microwave integrated circuit technology on RT/Duroid substrate (dielectric constant=10 2) The configuration of this coupler is

gıvcn ın figuıe 3 1 Thıs chapter shows that the desıgn data for the standard mıcıostrıp and slot lıne aıe good approxımatıons for dımensıonıng the coupler cross sectıon The accuı ıcy of the analysıs ın chapter2 ıs verıfied by a comparıson wıth the mcasuıement data



Γigure 3 1 Conﬁguratıon of the mıcrostrıp –slot coupler (a) Upper sıde of the substıate (b) Cıoss sectıon

In addıtıon desıgn parameters that wıll remaın to be determıned such as posıtıon of the reference planes The definıtıon of the slot lıne characterıstıc ımpedance and the

influence of the transmission line loss are investigated The improvement of the coupler performance attainable through compensation is shown with the help of theoretical analysis

3 2 Simplified computation of even and odd mode Parameters

The accurate comparison of the characteristic impedance Z_M and effective permittivity ε_{ee} for the microstrip mode (even –mode excitation) and the corresponding values Z_S and ε_{eo} for the slot mode (odd mode excitation) of the original coupler cross section shown in figure 3 1 requires an extensive numerical analysis In practice such data are often not available Analyses are however usually available for the standard microstrip transmission line These analyses can provide useful approximate parameter values for the microstrip slot configuration The errors in Z_M and ε_{ee} under the assumption that the slot width s is negligibly small for s/h<0 3 Coupling coefficient of ≤6 dB is taken into account here Thus the parameters for 3 dB microstrip – slot couplers for the conventional microwave frequency range (approximately 2 18GHZ) are covered by the approximations of the standard microstrip and slot line

3 3 Comparison of measured and theoretical Results

3 3 1 Implemented coupler

Microstrip slot coupler with a center frequency f_c in the region of 6 74024 GHZ is implemented on RT/Duroid substrate using thin film technology with 6 μm thick copper conductors view of which is shown in figure 3 2

The coupler is partially compensated This means that the length l_S of the compensation slot line is >0 But the compensation condition 2 5 3(h) of chapter 2 for l_S is not met, the coupler is said to be 3 dB coupler if it satisfies the matching condition $Zo = \sqrt{Z_L Z} = \sqrt{Z_M Z_S}$ But this type of coupler is not realizable because

of non-availability of correct thickness substrate, mask aligning technique. So the design corresponding to the available substrate is actually for 5.31- dB coupler with slot width 0.14 mm and center frequency 6.74024 GHZ.

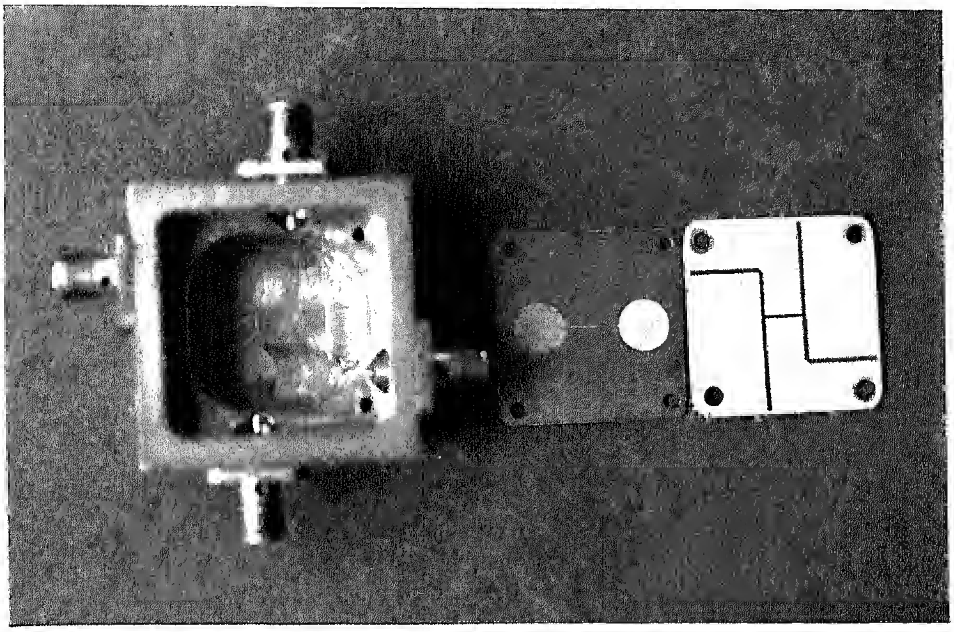


Figure 3.2 Photograph of the unassembled view of the microstrip-slot coupler

The measured frequency responses of the transmission loss $a_{21}= -20\log|S_{21}|$, coupling loss $a_{31}= -20\log|S_{31}|$ and isolation loss $a_{41} = -20\log|S_{41}|$ of the implemented coupler are compared with the theoretical responses. S_{ij} are the scattering parameters of the coupler. The measured frequency responses of a_{21} and a_{31} refer the whole circuit on the substrate shown in figure 3.1, and include the attenuation of 0.1 dB in all two feeding lines. The theoretical values of S_{21}, S_{31} were compared with the aid of the analysis given in chapter 2, with Z_0=50Ω

3.3.2 Empirical Determination of Reference Plane Location and of proper slot Line Impedance Definition

For the calculation of the theoretical responses of $a_{21}(f)$ and $a_{31}(f)$ for the coupler design in general two questions arise First at what position should the reference planes T_1 T_2 be placed? And second which definition should be chosen for the computing the characteristic impedance Z_S and Z_M? The answers can be obtained by comparing the theoretical responses with the measured frequency responses

For the first question the tentative positions of the reference planes described by the distance d between T_1 (orT_2) and outer edge of the conductor of the feeding microstrip line is d= $\frac{w_p}{2}$ and the corresponding value of the effective length l of the coupling section is shown

Here we have considering d=$\frac{w_l}{2}$ for the calculation center frequency corresponding to which $l = \frac{\lambda}{4} = 5\ 62$ mm

The given substrate thickness is 0 325 mm and ε_r=10 2 on which coupler fabricated consist microstrip line width w=0 42 mm and slot line width s=0 14 mm For microstrip line $\frac{w}{h} = 1\ 292$ and for slot case $\frac{s}{h} = 0\ 43076$

3 4 Approximate Calculations

Closed form expressions for characteristic impedance and slotline wavelength have been obtained by curve fitting the results based on the Cohn s analysis These expressions have an accuracy of about 2% for the following parameters

$9\ 7 \le \varepsilon_r \le 20$

$0\ 02 \le \frac{w}{h} \le 1\ 0$ and $0\ 01 \le \frac{h}{\lambda_o} \le \left(\frac{h}{\lambda}\right)$

Where $\left(\frac{h}{\lambda_c}\right)$ is the cut off value for the TE_{10} surface wave mode on slotline and is given by $\left(\frac{h}{\lambda}\right) = \frac{0\ 25}{\sqrt{\varepsilon_r - 1}}$

For $0\ 2 \leq \frac{s}{h} \leq 1\ 0$

$$\frac{\lambda_s}{\lambda} = 0\ 987 - 0\ 21 \ln \varepsilon\ + \frac{s}{h}(0\ 111 - 0\ 0022\varepsilon\ ) - \left(0\ 053 + 0\ 41\frac{s}{h} - 0\ 0014\varepsilon\ \right) \ln\left(\frac{h}{\lambda_0}\ 10^2\right)$$

3 4(a)

By substituting the value of ε_r and $\frac{s}{h}$ in above equation we get

$$\frac{\lambda_s}{\lambda} = 0\ 4794$$

And $\varepsilon_{co} = 5\ 711$

For $0\ 2 \leq \frac{s}{h} \leq 1\ 0$

$$Z_S = 113\ 1923\ 257 \ln \varepsilon_r + 1\ 25\frac{w}{h}(114\ 59 - 22\ 531 \ln \varepsilon\ ) + 20\left(\frac{w}{h} - 0\ 2\right)\left(1 - \frac{w}{h}\right)$$

$$\left[0\ 15 + 0\ 1 \ln \varepsilon_r\ + \frac{w}{h}(-0\ 79 + 0\ 899 \ln \varepsilon_r)\right]$$

$$\left\{\left[10\ 25 - 2\ 171 \ln \varepsilon_r\ + \frac{w}{h}(2\ 1 - 0\ 617 \ln \varepsilon_r) - \frac{h}{\lambda}\ 10^2\right]^2\right\} \qquad 3\ 4\ (b)$$

Putting the values of ε_r=10 2 $\frac{s}{h}$ = 43074 and $\frac{h}{\lambda} = \left(\frac{h}{\lambda}\right) = \frac{0\ 25}{\sqrt{\varepsilon_r - 1}} = 0\ 0825$ we get

Z_S=88 25 Ω

and $\varepsilon_{eo} = 5\ 711$

Similarly calculations for microstrip related parameters

Corresponding to $\frac{w}{h} = 1\ 292$ under quasi static approximations and assuming the strip thickness to be negligible the expression for Z_M is given and the accuracy of this expression is reported to be better than 0 2% for $\varepsilon_r \leq 128$ and $0\ 01 \leq \frac{w}{h} \leq 100$

$$\varepsilon \;=\left\{\frac{(\varepsilon\;+1)}{2}\right\}+\left\{\frac{(\varepsilon\;-1)}{2}\right\}\left\{1+\left(\frac{10}{2}h\right)\right\}^{al} \qquad 3\;4(c)$$

Where

$$a=\left[1+\left(\frac{1}{49}\right)\ln\left\{\frac{\left(\frac{w}{h}\right)^{4}+\left(\frac{w}{52h}\right)}{\left(\frac{w}{h}\right)^{1}+0\;432}\right\}+\left(\frac{1}{18\;7}\right)\ln\left\{1+\frac{w}{18\;1h^{3}}\right\}\right] \qquad 3\;4(d)$$

and

$$b=0\;564\left(\frac{\varepsilon\;-0\;9}{\varepsilon_{r}+3}\right)^{0\;053} \qquad 3\;4(e)$$

So after substituting the known parameters in the equation we get

$$\varepsilon_{r}\;=6\;944$$

$$Z_M=\left\{\frac{60}{\sqrt{\varepsilon}}\right\}\ln\left[\left(\frac{h}{w}\right)F\left(\frac{w}{h}\right)+\left\{1+\left(\frac{2h}{w}\right)^{2}\right\}^{\frac{1}{2}}\right] \qquad 3\;4(f)$$

Where

$$F\left(\frac{w}{h}\right)=\left[6+(2\pi-6)\exp\left\{-\left(30\;666\frac{h}{w}\right)^{0\;7528}\right\}\right] \qquad 3\;4(g)$$

So the value of $Z_M = 43\;36\;\Omega$

$$Z_0=\sqrt{Z_M\;Z_s}=61\;85\,\Omega$$

For center frequency

$$\frac{\lambda_g}{\lambda_o}=\frac{1}{\sqrt{\varepsilon_{ff}}}$$

λ_g From coupler structure =4×length from T_1 plane to T_2 plane =22 48 mm

Putting the numerical values of the parameters and calculating we get

f_c =6 7404 GHz

The increased characteristic impedance required for compensation

$$Z_S = \frac{Z_S}{2}\left[\cot\left(\frac{\pi}{4}\sqrt{\frac{\varepsilon}{\varepsilon}}\right)+\tan\left(\frac{\pi}{4}\sqrt{\frac{\varepsilon}{\varepsilon}}\right)\right] \qquad 3\,4(h)$$

$$= 89\ 35\Omega$$

The length of the compensation lines which is assumed to have approximate value is chosen according to the following expression

$$l_s = l\sqrt{\frac{\varepsilon}{\varepsilon}}\frac{2}{\pi}\arctan\left\{\frac{Z_s}{Z_S}\left[1-2\sin^2\left(\frac{\pi}{4}\sqrt{\frac{\varepsilon}{\varepsilon}}\right)\right]\right\} \qquad 3\,4(i)$$

$l_s = 0\ 58$ mm

The input susceptance obtained at the frequency f_c=6 74024 GHZ

$$B_C = \frac{1}{Z_s}\tan\left(\frac{2\pi l_s f\sqrt{\varepsilon}}{c}\right)$$

$B_C = 3\ 632\times10^{3}$ mho

So computed data which are relevant to realized coupler are as follows

Dielectric constant of the substrate (RT duroid)	$\varepsilon_r = 10\ 2$
Thickness of the substrate	h = 0 325 mm
Width of the microstrip line	w = 0 42 mm
Width of the slot line	s = 0 14 mm
Width of the feed lines	w_p = 0 62 mm
Length of the microstrip line	l = 5 62 mm
Compensation length of the slot	l_s = 0 58 mm
Effective permittivity of the microstrip medium	ε_{ee} = 6 944
Effective permittivity of the slot line medium	ε_{eo} = 5 711
Slotline characteristic impedance	Z_S = 88 25 Ω
Microstrip line characteristic impedance	Z_M = 43 36 Ω
Characteristic impedance of the coupler	Z_O = 61 85 Ω
Effective permittivity of the coupled medium	ε_{eff} = 6 32
Center frequency	f_c = 6 7404 GHz

Input susceptance corresponding to center frequency Bc = 3 6× 10^{3} seimens

Since Bc is shunted across the ends of the coupling section So the reflection coefficients for odd mode excitation can be computed from the input susceptance of the transmission lines shunted with the susceptance 2Bc

$$\frac{\theta}{2} = \omega\, l \sqrt{\frac{\varepsilon}{c}} = 1\ 0886 \text{ radian}$$

$$\frac{\theta}{2} = \omega\, l \sqrt{\frac{\varepsilon}{c}} = 1\ 204 \text{ radian}$$

By putting the values of the parameters and calculating we get the input reflection coefficients for the compensated coupler as follows

$$\Gamma_{\iota} = \exp\left\{-j2\arctan\left[\frac{2Z}{Z_s}\tan\left(\frac{\theta}{2}\right)+2B_c Z\right]\right\} \Rightarrow \exp(j2\ 3873) \qquad 3\ 4(j)$$

$$\Gamma = \exp\left\{-j\arctan\left[-\frac{2Z}{Z_s}\cot\left(\frac{\theta}{2}\right)+2B_c Z\right]\right\} \Rightarrow \exp(j0\ 4534) \qquad 3\ 4(k)$$

$$\Gamma_{\prime\prime} = \exp\left\{-j2\arctan\left[\frac{Z}{2Z_M}\tan\left(\frac{\theta}{2}\right)\right]\right\} \Rightarrow \exp(j1\ 956) \qquad 3\ 4(l)$$

$$\Gamma_{\iota} = \exp\left\{j2\arctan\left[\frac{Z}{2Z_M}\cot\left(\frac{\theta}{2}\right)\right]\right\} \Rightarrow \exp(j\ 4410) \qquad 3\ 4(m)$$

For these input reflection coefficients we get

Return loss $= 20\log|S_{11}| \Rightarrow 9\ 74$ dB 3 4(n)

Coupled port output $= 20\log|S_{31}| \Rightarrow 5\ 31$ dB 3 4(o)

Direct port output $= 20\log|S_{21}| \Rightarrow 2\ 36$ dB 3 4(p)

Isolated port output $= 20\log|S_{41}| \Rightarrow 19\ 6$ dB 3 4(q)

Directivity $= 20\log\left|\frac{S41}{S31}\right| \Rightarrow 14\ 29$ dB 3 4(r)

For the second question three definitions were investigated for Z_S and Z_M $Z^{(PV)}=V^2/(2P)$ $Z^{(VI)}=V/I$ and $Z^{(PI)}=2P/I^2$ where P denotes the power transported in the direction of propagation of the mode I denotes the longitudinal

current on one of the two electrodes of the slot line or the strip conductor of the microstrip transmission line and V denotes the voltage over the slot in the slot line or the voltage between the center of the strip conductor of the microstrip line and the ground The conversion formula is

$$Z^{(II)}=[Z^{(VI)}]^2/Z^{(PV)}$$

Linear interpolation is possible with good approximation (error<0 5 percent) between 4GHZ and 6GHZ and between 6GHZ and 8GHZ

We first consider the 3 dB coupler which is matched to Z_O =50Ω the four positions of the reference planes T_1 and T_2 were used to calculate the theoretical values of a_{21} (f) and a_{31} (f) It can be seen that the closest agreement with the measured frequency responses occurs for the reference plane positions defined by $d=\frac{w_l}{2}$ which can be alternatively expressed by l=a+$\frac{w_p}{2}$ Thus T_1 and T_2 are located approximately $w_p/2$ within the outer edges of the feeding lines of width W_P Larger value of l result in frequency responses that deviate more sharply from the measurements

The theoretical responses for a_{21} (f) and a_{31} (f) were calculated using the one Z_S definition the position $d=\frac{w_l}{2}$ of the reference planes For the microstrip mode we used the definition Z_M The difference between the Z_M definitions is below 5 percent and in the first approximation negligible Also negligible is the terminal capacitance of the slot line open circuits When Z_S definition is used the theoretical responses for a_{21} (f) should almost coincides with the measured response of the coupling section without feeding lines, which is obtained by subtracting the 0 1dB attenuation to the feeding lines in our case there is deviation because of alignment problem As slot is not completely beneath the microstrip line on the other side of the substrate Perfect alignment is only possible if alignment is done with mask aligner Although theory should yield lower values as the dissipative losses are not being taken in to account The computed values of a_{21} (f) and a_{31} (f) are average of 0 4 dB below the measured

values This diffeience is attiibuted due to the ohmic dielectric and the radiation losses at discontinuities that aie not taken in to account in the theory Thus the definition of Z_S is the moie pieferable one for the coupler design

The choice of the iefeience plane positions and the slot line impedance definition obtained foi the matched coupler is also applies to coupler having slot width s=0 14mm

As the accuiacy of the cuives concerns the following information should be added heie Measuiement of the substiate thickness h over the whole substrate iesulted in maximum deviation of 8µm from the nominal value of 0 325mm which keeps the change in Z_S and Z_M below 0 5 peicent and that of the ε_{eo} and ε_{ee} 0 1 peicent also the dielectric constant tolerance $\Delta\varepsilon r/\varepsilon r$<3percent results in a change of <2 peicent foi Z_S Z_M εeo and ε_{ee} Thus the total maximum error for Z_S and Z_M of 2 5 percent is noticeably smaller than the typical difference of 8 to 20 percent between the vaiious Z_S definitions Similarly the possible frequency shift of <1 peicent foi the theoietical results due to tolerances in ε_r and h is noticeable smaller than the centei frequency shift of 8 to 30 percent caused by different reference plane positions d= 0 to 1 5 w_p The measurements were performed using PC interfaced powei meter directional coupler and phase locked frequency synthesizer

3 5 Measurements and results of the microstrip-slot coupler

Measurements aie done by netwoik analyzer made out by interconnecting power metei direction couplei and frequency synthesizer (2 20 GHz) and its operation controlled by computer Only the magnitude of S_{11} S_{21} $S_{31}$and S_{41} of the device are characteriзed S_{11} S_{21} $S_{31}$and S_{41} represent the reflection Transmission coupling and isolation parameters respectively Figure 2 1 illustrates the microstrip slot coupler It can be seen that the measured center frequency is in good approximation with the theoretical value 6 70404 GHz The combined plot of experimental results taken is given in figure 3 3

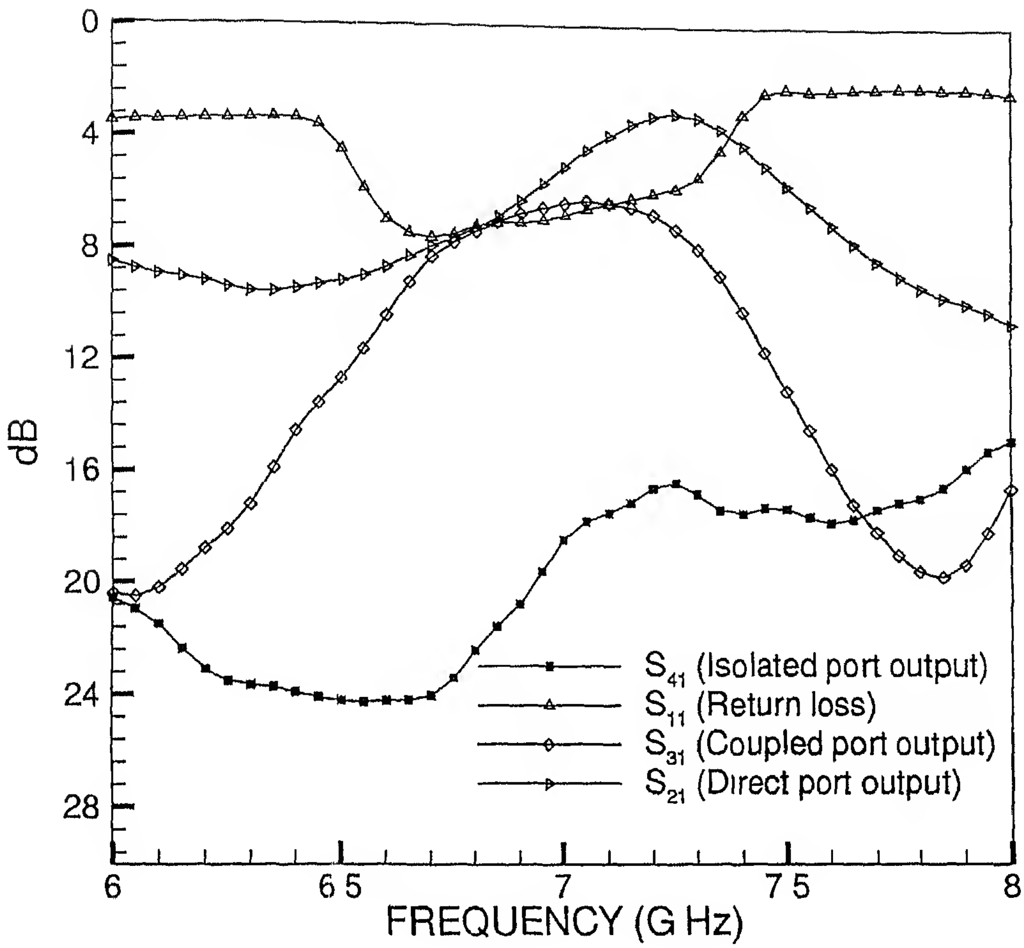


Figure 3 3 Plot showing measured magnitude of S_{41} S_{11} S_{31} and S_{21} of microstrip slot coupler

Specifications of coupler

Coupling = 6 dB

Return loss = 7 5 dB

Coupling band width = 15%

Isolation band width = Better than 15%

Isolation = Better than – 20 dB

Center frequency 7 0 GHz

The isolation is better than the theoretical value 19 6 dB but direct port output 3 5 dB is fairly in good agreement over the limited range of frequencies with the

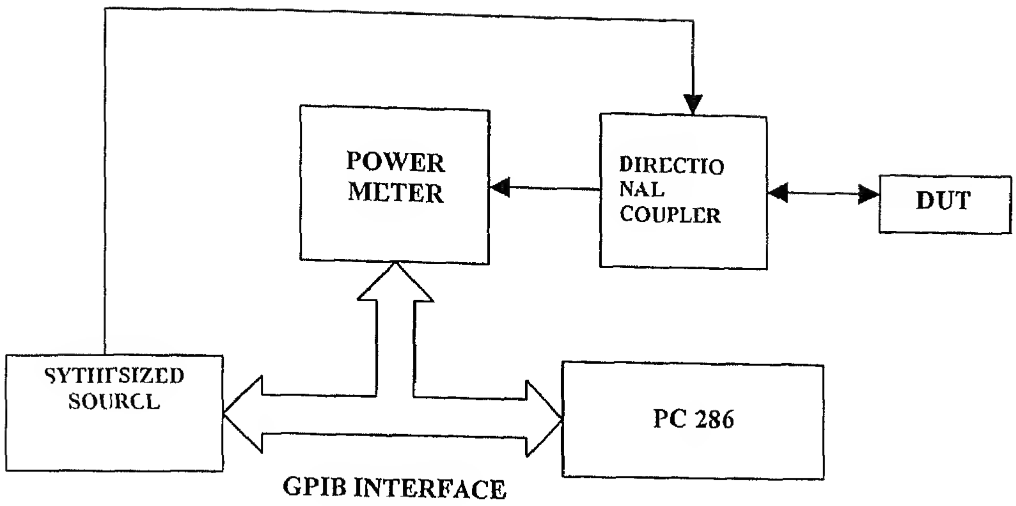


Fig 3 4 Block diagram of the experimental setup for return loss measurement

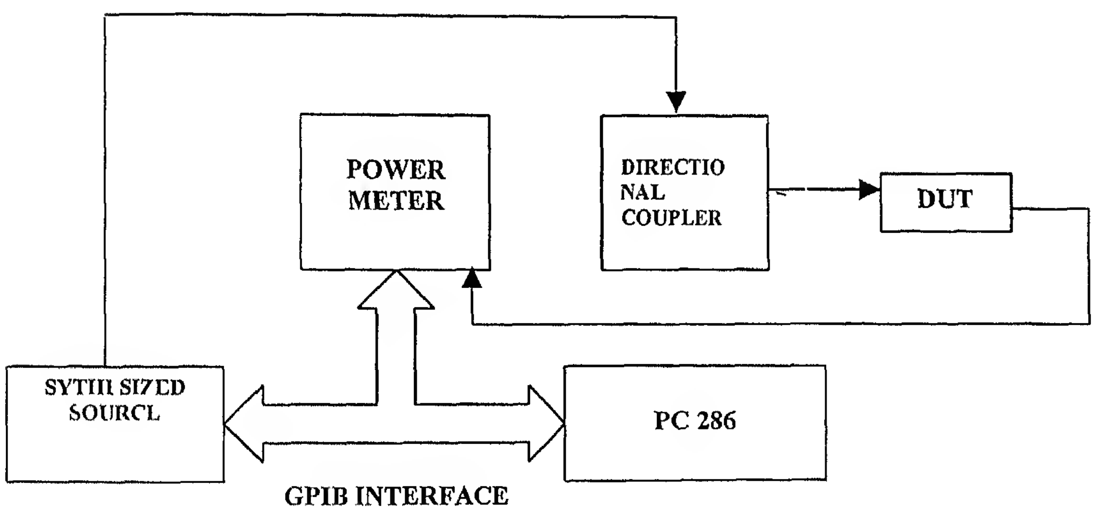


Fig 3 5 Block diagram of the experimental setup for direct port coupled port and isolated port output

The isolation is better than the theoretical value 19 6 dB but direct port output 3 5 dB is fairly in good agreement over the limited range of frequencies with the theoretical predicted value 2 36 dB Coupling from 6 5 GHz – 7 5 GHz is in the range 6dB to 7 dB which is not in good agreement with the approximate theoretical predicted value 5 31 dB It can also be seen from the plot that return loss is 7 5 dB which is not in good agreement with the theoretical predicted value 9 74 dB in the band of interest This is due to mismatch and transition losses at connection ports

CHAPTER 4

Summary and conclusion

4 1 Summary

A novel method of analysing microstrip slot coupler theory and realization of the same had been described in this thesis The microstrip slot coupler is special classes of couplers for that type of coupler expression of scattering parameters are derived in terms input reflection coefficients by using even odd mode analysis of a reciprocal passive linear four port network with double symmetry with reference to the two symmetry planes Derivation of the scattering parameters is done in both the cases i e without compensation and with compensation lines Compensation is accomplished by lengthening the slot line to capacitive loading at the ends of the coupled region the isolation has been improved by 5 10 dB as compared to the uncompensated structure The frequency response of the isolation depends on the ratio of $\frac{l_s}{Z_s}$ Special cases of microstrip slot couplers retaining ideal theoretical approach can not be realized with the real coupler configuration because in this case $\varepsilon < \varepsilon_{ee}$ however it can be used to derive design equations for the real couplers The ideal coupler is characterized by the reflection coefficient S_{11} being zero and S_{41} being zero i e infinite directivity at all frequencies For real uncompensated commercial coupler of conventional dimension $l = 0$ $\varepsilon_{ee} > \varepsilon_{eo}$ and $v_o > v_e$ The real couplers with compensation ($\varepsilon_{eo} < \varepsilon_{ee}$ $l > 0$) and matching i e $S_{41}=0$ $S_{11}=0$ can be realized with the aid of the compensation lines S only at a single arbitrary chosen frequency viz the compensation frequency f_{co} for that it is essential that matching condition should be satisfied

Practical designing aspects of the microstrip slot coupler on RT/ Duroid substrate with compensation are treated in the supplementation of the theoretical analysis Emphasis is given to the realization of 4 5 d B coupler on the 0 325 mm thick

substrate with $\varepsilon_r = 10.2$ substrate. We had design data for the realization 4.5 dB coupler on the Al_2O_3 substrate with relative permittivity of 9.8 and substrate thickness 0.635 mm. But due to non availability of the compatible substrate the fabrication is done on the substrate having thickness 0.325 mm for the slotline 0.14 mm on the ground plane and microstrip width 0.42 mm on the other plane for the mismatch condition. The theoretical coupling is 5.31 dB for the given substrate having same dimensions of slot line and microstrip as was implemented on the substrate of thickness 0.625 mm. The coupler is working satisfactory with center frequency 7.0 GHz. The practical coupling value is not equal to the approximately calculated value because of the slot is not accurately beneath the microstrip in the ground plane, this is because of the non availability of the mask aligning facility. If the alignment is done with the mask aligner we can get coupler of which coupling has close agreement with the calculated values for the perfect matching condition. For the isolation improvement compensation is added in this type of coupler which has removed the junction effect with enhancing the directivity. In uncompensated type of coupler such a good directivity can not be obtained. In this thesis comparison has been made for the theoretical and practical parameters.

4.2 Conclusion

Scattering parameters of the coupler are derived in this thesis by using the even odd mode analysis for four port network having double symmetry, equations have been derived for the frequency dependent scattering parameters S_{ij} of the microstrip slot coupler, whereby different phase velocities v_e, v_o of the even and odd modes, as well as the mismatch, are included through the arbitrary choice of the characteristic impedance Z_0 and the even mode and odd mode characteristic impedances Z_e, Z_o. The analysis was extended to cover the coupler with supplementary slot lines as conventionally used for compensation. For $v_e = v_o$ the parameters of an ideal TEM coupler are realized. Simple design specifications are derived for the compensation slot lines.

Comparison of the measurement of coupler with a slot width 0 14 mm realized on RT/ Duroid substrate with results yielded by the theoretical analysis show close agreement under the following condition First the reference planes for the ends of the coupling section must lie within about half width of the feeding lines with center frequency in the region of 7 GHz Second the characteristic impedance Z_S defined by the current and voltage must be used for the slot line Third in order to take of the transmission line loss the theoretically calculated parameters must be displaced by a fixed value – here approximately 0 4 dB – towards higher attenuations Estimates have shown the theories of the standard microstrip transmission line and the standard slot line to be sufficient for the practical designing of the coupler design

4 3 Scope for future work

The microstrip slot line coupler can be integrated on the high relative permittivity substrate to improve the dispersion loss of the coupler as effective permittivity in slotline propagation mode becomes much lower in comparison to the relative permittivity of the used substrate If available the intregation of the coupler can be done on the transparent substrate having high relative permittivity of the order 20 to 30 which will eliminate the error occurring in the alignment of the microstripline and slotline If the transparent substrate is not available the alignment requires mask alignment technique during fabrication process For accurate theoretical calculation of parameters full wave analysis can be used with which we can get design parameters very close to practical values of the implemented coupler

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